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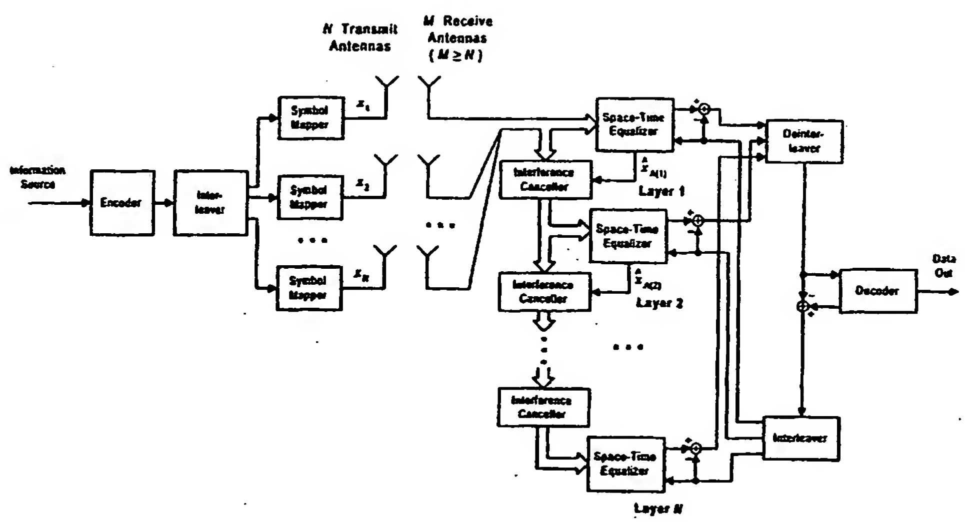
(54) Title: TURBO DETECTION OF SPACE-TIME CODES



Coded layered space-time architecture: LST-I.

(57) Abstract: Communication systems which employ multiple transmit and receive antenna-element arrays. Data streams for transmission may be interleaved among the transmit antenna elements in order to reduce decision errors. Turbo processing of equalizer output from a number of layers in a layered space-time processing architecture may be employed to reduce decision errors. Additionally, space-time equalization may be performed to maximize signal to noise ratio such as via minimum mean square error processing, rather than zero forcing, in order to achieve the Shannon limit, reduce multi-path effects and/or reduce intersymbol interference. Moreover, the receiver can select number and/or identity of receive antenna elements from among a larger group in order to optimize performance of the system.

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TURBO DETECTION OF SPACE-TIME CODES

RELATED APPLICATIONS

This document claims priority to and incorporates by reference copending provisional USSN 60 / 152,982 entitled "Turbo Space-Time Processing to Improve Wireless Channel Capacity" filed on September 9, 1999.

FIELD OF INVENTION

The present invention relates to systems and processes for radio communications using multiple-element antenna array technology.

BACKGROUND

Turbo processing and space-time equalization are terms that comprehend several conventional ways to increase wireless channel capacity. Generally, turbo coding and/or processing refers to techniques aimed at approaching the Shannon limit in a channel, while space-time processing refers to techniques for processing signals from multi-element antenna arrays to exploit the multi-path nature of fading wireless environments.

European patent application no. EP 817 401 A2 published July 1, 1998 in the name of Foschini discloses the use of a number of processing layers for space time processing of signals from multiple-receiver antenna elements. There, the transmitter feeds a number of transmitter antenna elements by cyclically apportioning segments of the modulated encoded stream of data to transmitter antenna elements. At the receiver, a number of receiver antenna elements are coupled to a number of processing layers, in order to perform the space-time processing. Signal components received during respective periods of time over a plurality of the receive antenna elements are formed into respective space and time relationships in which space is associated with respective transmitter antenna elements. Preprocessing occurs so that a collection of signal components having the same space-time relationship

forms a signal vector such that particular decoded signal contributions can be subtracted from the signal vector while particular undecoded contributions can be nulled out of the signal vector. The resulting vector is then supplied to a decoder for decoding to reform the data stream. Such conventional systems and techniques are further described in documents referred to in the "Detailed Description" section of this document.

SUMMARY OF THE INVENTION

Systems and processes according to the present invention employ a number of transmitter antenna elements and a number of receiver antenna elements coupled to multiple space-time processing layers in the receiver. In the present invention, however, portions of the information stream being communicated can be interleaved among transmitter antenna elements such as on a random or pseudo random basis; among other things, such interleaving decreases decision errors in the space-time equalization process. Furthermore, each processing layer preferably includes turbo processing in order to feed soft decisions about information being processed back to the equalizers. Moreover, space-time equalization processes according to the present invention preferably seek to maximize signal to noise ratio rather than zero forcing, as well as reduce multi-path effects and intersymbol interference. A preferred process uses minimum mean square error processing which allows the Shannon limit actually to be achieved. Furthermore, systems and processes according to the present invention preferably allow selection of the number and identity of receiver antenna elements to which the receiver may be coupled in order to optimize performance.

According to one embodiment of the invention, an information source is coupled to provide a plurality of data streams to a plurality of transmit antennas, via, for each stream, an encoder, interleaver and symbol mapper. On the receiver side, a plurality of M receiver elements are coupled to a plurality of processing layers. The number of receiver antenna elements M is preferably greater than or equal to the number N of transmit antenna elements, since equalization according to the present invention does not

require an extra degree of freedom. The M receiver antenna elements are coupled to the first processing layer by coupling to a space-time equalizer which preferably applies minimum mean square error processing in order to maximize signal to noise ratio. The output of the equalizer is applied to a deinterleaver, after which the deinterleaved stream is supplied to a decoder in the layer. Output of the decoder is provided for output common with the output from the other decoders in the other layers. Preferably, each layer also includes an interleaver which receives output from the decoder and deinterleaver and supplies its interleaved output back to the equalizer in the layer in order to provide soft decision making to the equalizer. In successive processing layers, output from the decoder of the preceding layer is combined with information from the interference canceler of the layer preceding the preceding layer (except the second layer, which receives signals from an interference canceller which is coupled to the decoder of the first layer and to the receive antenna elements).

According to an alternate embodiment, the deinterleaver, interleaver and decoder are shared among layers, so that the equalizer of each layer outputs to a deinterleaver common to all layers. The output of the deinterleaver may then be coupled to a decoder which again is common to all layers. An interleaver may be provided which receives output from the deinterleaver and the decoder and applies it to each equalizer for soft decisions to be applied to the equalizers.

Accordingly, components for deinterleaving, decoding and reinterleaving may be functionally located in each layer, or common to the layers. In the first case, each layer below the first layer processes signals from an interference canceller which receives signals from a decoder in the preceding layer and from the antenna elements (in the case of the second layer) or the interference canceller in the next-preceding layer (in the case of other layers). In the second case, each layer below the first processes signals from an interference canceller which receives signals from the equalizer in the preceding layer and from the antenna elements (in the case of

the second layer) or the interference canceller in the next preceding layer (in the case of other layers). Such turbo processing architectures can be used in connection with layered space-time equalization which relies on zero forcing rather than minimum square error processing. They can also be used in multi array systems in which the data streams are periodically cycled rather than interleaved.

It is accordingly an object of the present invention to provide improved layer space-time processing for communication systems which employ turbo processing techniques in order, among other things, to reduce decision errors.

It is an additional object of the present invention to provide layered space-time processing for communication systems which seeks to maximize signal to noise ratio, thereby better addressing the Shannon limit, and which also addresses mulit-path effects and / or intersymbol interference.

It is an additional object of the present invention to provide processing for communication systems in which data streams may be interleaved rather than periodically cycled among transmit antenna elements, in order, among other things, to reduce decision errors.

It is an additional object of the present invention to provide layered space-time processing for communication systems in which a receiver can select a set of antenna elements, including their number and / or identity, from among a larger group of antenna elements in order to optimize performance of the system.

Other objects, features, and advantages of present invention will become apparent with respect to the remainder of this document.

BRIEF DESCRIPTION OF THE DRAWINGS

Fig. 1(a) is a schematic diagram showing a first embodiment of communications systems according to the present invention.

Fig. 1(b) is a schematic diagram showing a second embodiment of communications systems according to the present invention.

Fig. 2 is schematic diagram showing one form of space-time processing according to the present invention.

Figs. 3(a) and 3(b) are diagrams which compare performance between two coding schemes according to the present invention.

Fig. 4 is a diagram which shows different capacity bounds for processing according to the present invention over a flat Rayleigh fading channel.

Fig. 5 is a diagram which shows simulation results for a system according to the first embodiment of the present invention with two transmit and two receive antenna elements.

Fig. 6 is a diagram which shows simulation results for a system according to the first embodiment of the present invention with two transmit and four receive antenna elements.

Fig. 7 is a diagram which shows simulation results for a system according to the first embodiment of the present invention with two, four and eight receive antenna elements, using soft decisions and six turbo iterations.

Fig. 8 is a diagram which shows simulation results for a system according to the second embodiment of the present invention with two, four and eight receive antenna elements, using soft decisions and two turbo iterations.

Figs. 9(a) and 9(b) are diagrams which show simulated performance of the first embodiment of the present invention using soft decisions and four antenna elements for typical urban and hilly terrain profiles.

DETAILED DESCRIPTION

The documents and references cited in the following disclosure are incorporated herein by this reference.

Abstract—By deriving a generalized Shannon capacity formula for multiple-input, multiple-output Rayleigh fading channels, and by suggesting a layered space–time architecture concept that attains a tight lower bound on the capacity achievable, Foschini has shown a potential enormous increase in the information capacity of a wireless system employing multiple-element antenna arrays at both the transmitter and receiver. The layered space–time architecture allows signal processing complexity to grow linearly, rather than exponentially, with the promised capacity increase. This paper includes two important contributions: First, we show that *Foschini's lower bound is, in fact, the Shannon bound* when the output signal-to-noise ratio (SNR) of the space–time processing in each layer is represented by the corresponding "matched filter" bound. This proves the optimality of the layered space–time concept. Second, we present an embodiment of this concept for a coded system operating at a low average SNR and in the presence of possible intersymbol interference. This embodiment utilizes the already advanced space–time filtering, coding and turbo processing techniques to provide yet a practical solution to the processing needed. Performance results are provided for quasi-static Rayleigh fading channels with no channel estimation errors. We see for the first time that the Shannon capacity for wireless communications can be both *increased* by N times (where N is the number of the antenna elements at the transmitter and receiver) and *achieved* within about 3 dB in average SNR, about 2 dB of which is a loss due to the practical coding scheme we assume—the layered space–time processing itself is nearly information-lossless!

Index Terms—Equalization, interference suppression, space–time processing, turbo processing.

I. Introduction

"TURBO" and "space–time" are two of the most explored concepts in modem-day communication theory and wireless research. From a communication theorist's viewpoint, "turbo" coding/processing is a way to approach the Shannon limit on channel capacity, while "space–time" processing is a way to increase the possible capacity by exploiting the rich multipath nature of fading wireless environments. We will see through a specific embodiment in this paper that combining the two concepts provides even a practical way to both increase and approach the possible wireless channel capacity.

With growing bit rate demand in wireless communications, it is especially important to use the spectral resource efficiently. The basic information theory results reported by Foschini and Gans [1] have promised extremely high spectral efficiencies possible through multiple-element antenna array technology. In high scattering wireless environments (e.g., troposcatter, cellular, and indoor radio), the use of multiple spatially separated and/or differently polarized antennas at the receiver has been very effective in providing diversity against fading [2], [3]. Receiver diversity techniques also create signal processing opportunities for interference suppression and equalization (e.g., [4]–[6]). However, using multiple antennas at either the transmitter or the receiver does not enable a significant gain in the possible channel capacity. According to [1], the Shannon capacity for a system with 1 transmit and N receive antennas scales only logarithmically with N, as $N \to \infty$. For a system using N transmit and 1 receive antennas, asymptotically there is no additional capacity to be gained, assuming that the transmit power is divided equally among the N antennas.

Foschini and Gans [1] have shown that the asymptotic capacity of multiple-input, multiple-output (MIMO) Rayleigh fading channels grows, instead, linearly with N when N antennas are used at *both* the transmitter *and* the receiver. Furthermore, in [7], Foschini suggested a layered space–time architecture concept that can attain a tight lower bound on the capacity achievable. In this layered space–time architecture, N information bit streams are transmitted simultaneously (in the same frequency band) using N diversity antennas. The receiver uses another N diversity antennas to decouple and detect the N transmitted signals, one signal at a time. The decoupling process in each of the N processing "layers" involves a combination of *nulling out* the interference from yet undetected signals (N diversity antennas can null up to $N - 1$ interferers, regardless of the angles-of-arrival [5]) and *canceling out* the interference from already detected signals. One very significant aspect of this architecture is that it allows an N-dimensional signal processing problem—which would otherwise be solvable only through multiuser detection methods [8] with m^N complexity (m is the signal constellation size)—to be solved with only N similar 1-D processing steps. Namely, the processing complexity grows only linearly with the promised capacity.

This paper includes two important contributions. First, we show that Foschini's lower bound is, in fact, the Shannon bound when the output SNR of the space–time processing in each layer is represented by the corresponding "matched filter" bound [6], i.e., the maximum SNR achievable in a hypothetical situation where the array processing weights to suppress the remaining interference in each layer are chosen to maximize the output signal-to-interference-plus-noise ratio and any possible

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intersymbol interference (ISI) is assumed to be completely eliminated by some means of equalization. The "matched filter" bound has been shown to be approachable using minimum mean-square error (MMSE) space–time filtering techniques [6].[1] By showing the equivalence of the generalized Foschini's bound and the Shannon bound, we essentially prove the optimality of the layered space–time concept.

Second, we present an embodiment of Foschini's layered space–time concept for a coded system operating at a low average SNR and in the presence of unavoidable ISI. Previously, a different embodiment has been provided in [9] for an uncoded system with variable signal constellation sizes, operating at a high average SNR without ISI. Adding coding redundancy might, at first, seem conflicting with the desire to increase the channel bit rate. Our justification is as follows: First, we seek to enhance the channel capacity from a system perspective. We use "noise" in SNR to represent all system impairments, including thermal noise and multiuser interference. The ability to operate at low SNR's means that more users per unit area can occupy the same bandwidth simultaneously. Second, we anticipate the use of adaptive-rate coding schemes to permit different degrees of error protection according to the channel SNR's. Incremental redundancy transmission [10], currently being considered for the Enhance Data Services for GSM Evolution (EDGE; GSM stands for Global System for Mobile Communications) standard, is an efficient way to implement adaptive code rates without requiring channel SNR monitoring. With such adaptive-rate coding, the system does not "waste" spectral resources under good channel conditions.

Meanwhile, the iterative processing principle used in turbo and serial concatenated coding [11]–[15] has been successfully applied to a wide variety of joint detection and decoding problems. One such application is the so-called "turbo equalization" [16]–[19], where successive maximum *a posteriori* (MAP) processing is performed by the equalizer and channel decoder to provide *a priori* information about the transmit sequence to one another. Similar to the layered space–time concept, turbo processing allows a multi-dimensional (*two*-dimensional in this case) problem to be optimally solved with successive 1-D processing steps without much performance penalty. In this paper, we apply the turbo principle to layered space–time processing in order to prevent decision errors produced in each layer from catastrophically affecting the signal detection in subsequent layers.

We consider two possible coded layered space–time structures: one applying coding across the multiple signal processing layers, and the other assuming independent coding within each layer. Similar to [1], we assume a *quasi-static* random Rayleigh channel model, where the channel characteristics are stationary within each data block, but statistically independent between different data blocks, different antennas, and, in the case of dispersive multipath channels, different paths. The system is assumed to have similar ISI situations as in EDGE and GSM, where multipath dispersions may last up to several symbol periods [20]. We show that near-capacity performance is achievable using 1-D processing and coding techniques that are already practical and "legacy-compatible" with the EDGE standard, e.g., the use of bit-interleaved 8-ary phase-shift keying (8-PSK) with rate-1/3 convolutional coding and an equalizer with a similar length and structure.

A slightly different layered space–time approach based on *space–time coding* [23], [24] has been studied in [25]. Although it is difficult to make a general comparison, we will see later that our coded layered space–time approach does by far outperform the results reported in [25] for $N = 4$ and $N = 8$. On the other hand, for $N = 2$, space–time coded quaternary phase-shift keying (QPSK) without layered processing appears to be the best known technique for achieving a spectral efficiency of 2 bps/Hz.

This paper is organized as follows. Section II provides a brief review of Foschini's layered space–time concept. Section III describes the two coded layered space–time architectures and presents a capacity analysis which reveals the equivalence of a generalized Foschini's lower bound formula and the true capacity bound. Section IV provides details on the array processing, equalization, and iterative MAP techniques. Section V presents performance results. A summary and conclusions are given in Section VI.

II. Background Theory

We briefly review the theory behind Foschini's layered space–time concept. The generalized Shannon capacity for a MIMO Rayleigh fading system with N transmit and M receive antennas is given in [1] as

$$C = \log_2 \left[\det \left(I + \frac{\rho}{N} HH^{\dagger}\right)\right] \tag{1}$$

where H is an $M \times N$ matrix, the (i, j)th element of which is the normalized channel transfer function of the transmission link between the jth transmit antenna and the ith receive antenna. I is the $M \times M$ identity matrix, ρ is the average SNR per receive antenna, and $\det(\cdot)$ and superscript $\dagger$ denote determinant and conjugate transpose. It is assumed that the transmit power is equally divided among the N transmit antennas. The normalization of the channel transfer function is done such that the average (over Rayleigh fading) of its squared magnitude is equal to unity.

The lower bound on capacity is provided in [1] as

$$C > \sum_{k=N-M+1}^{N} \log_2 \left[1 + \frac{\rho}{N} \chi_{2k}^2\right] \triangleq C_F \tag{2}$$

where χ_{2k}^2 is a chi-squared random variable with $2k$ degrees of freedom. For $M = N$

$$C_F = \sum_{k=1}^{N} \log_2 \left[1 + \frac{\rho}{N} \chi_{2k}^2\right]. \tag{3}$$

Since χ_{2k}^2 represents a fading channel with a diversity order of k, the lower-bound capacity in (3) can be viewed as the sum of the capacities of N independent channels with increasing diversity orders from 1 to N. This suggests a layered space–time approach [7] for detecting the N transmitted signals as follows:

[1] In a flat fading case, MMSE array processing achieves exactly the "matched filter" bound performance.

In the first layer, the receiver detects a first transmitted signal by nulling out interference from $N-1$ other transmitted signals through array processing. Assuming a "zero forcing" (ZF) constraint, one receive antenna is needed to completely correlate and subtract each interference [5]. Thus, the overall process of nulling $N-1$ interferences leaves the receiver with $N-(N-1)=1$ degree of freedom to provide diversity for detecting the first signal, i.e., a diversity order of 1 (or simply no diversity). Once detected, the first signal is subtracted out from the received signals on all N antennas.

In the second layer, the receiver performs similar interference nulling to detect a second transmitted signal. This time, since there are only $N-2$ remaining interferences, the receiver affords a diversity order of 2. The detected signal is again subtracted out from the received signals provided by the first layer.

Repeating the above interference nulling/canceling step through N layers, we see that the receiver affords an increasing order of diversity from 1 to N. If the capacities achieved in individual layers can be combined in some manner, then the layered space–time approach just mentioned will achieve the capacity lower bound expressed in (3). We will explore two capacity combining possibilities in the next section.

Note that the capacity and capacity low bound given in (1)–(3) are actually frequency-dependent. We here provide an explicit capacity formula for *band-limited, frequency-selective* channels (some variables are redefined to be consistent with later analytical development).

$$C = \langle \log_2[\det(\mathfrak{R}\aleph^{-1})] \rangle \tag{4}$$

where, as shown in equations (5)–(8) at the bottom of the page, $\mathfrak{R}$ is the frequency-domain correlation matrix of the signals on M receive antennas, $\aleph_j(f)$ is the noise power density at frequency f on the jth receive antenna, T is the symbol period, $H_{ij}(f)$ is the channel transfer function (not normalized) of the transmission link between the ith transmit antenna and the jth receive antenna, and superscripts $*$ and T denote complex conjugate and transpose. Note in (7) and (8) that we consider the folded spectra $H_{ij}(f-(m/T))$ and $\aleph_j(f-(m/T))$ of the channel transfer function and noise power density, where $m=-J,\ldots,J$ (J is finite because the signal sources are assumed to be band-limited). This is to take into account the effect of excess bandwidth and symbol-rate sampling when the frequency selectivity of the channel is not symmetrical around the Nyquist band edges. Even though we assume white Gaussian noise, the noise power density near and outside the Nyquist band edges actually attenuates with the receive filter transfer function. From our experiment (assuming a square-root Nyquist filter with a 50% rolloff factor), the computed capacity can be underestimated by as much as 0.5 dB if this attenuation is not taken into account.

III. Coded Layered Space–Time Architectures

A. Basic Concepts

We consider two coded layered space–time approaches as shown in Fig. 1(a) and (b). In the first approach, named "LST-I" (LST stands for "layered space–time"), the coded information bits are interleaved across the N parallel data streams $\boldsymbol{x}_1, \boldsymbol{x}_2, \cdots, \boldsymbol{x}_N$, where $\boldsymbol{x}_i$ denotes a sequence of complex-valued, transmit data symbols (e.g., 8-PSK symbols). The receiver first decouples the N data streams through interference nulling/cancellation, as described in Section II, then deinterleaves and decodes all the N data streams as one information block. In the second approach, "LST-II," the information is first divided into N uncoded bit sequences $\boldsymbol{u}_1, \boldsymbol{u}_2, \ldots, \boldsymbol{u}_N$, each of which is independently encoded, interleaved, and symbol-mapped to generate one of the N parallel data streams. At the receiver, the

$$\langle \cdot \rangle \triangleq T \int_{-(1/2T)}^{(1/2T)} [\cdot]\, df \tag{5}$$

$$\mathfrak{R} \triangleq \sum_{i=1}^{N} \boldsymbol{H}_i^* \boldsymbol{H}_i^T + \aleph \tag{6}$$

$$\aleph \triangleq \begin{bmatrix} \aleph_1\left(f-\frac{J}{T}\right) & & & & & \\ & \ddots & & & \mathbf{0} & \\ & & \aleph_M\left(f-\frac{J}{T}\right) & & & \\ & & & \ddots & & \\ & & & & \aleph_1\left(f+\frac{J}{T}\right) & \\ & \mathbf{0} & & & & \ddots \\ & & & & & \aleph_M\left(f+\frac{J}{T}\right) \end{bmatrix} \tag{7}$$

$$\boldsymbol{H}_i \triangleq \left[H_{i1}\left(f-\frac{J}{T}\right) \cdots H_{iM}\left(f-\frac{J}{T}\right) \cdots\cdots H_{i1}\left(f+\frac{J}{T}\right) \cdots H_{iM}\left(f+\frac{J}{T}\right)\right]^T \tag{8}$$

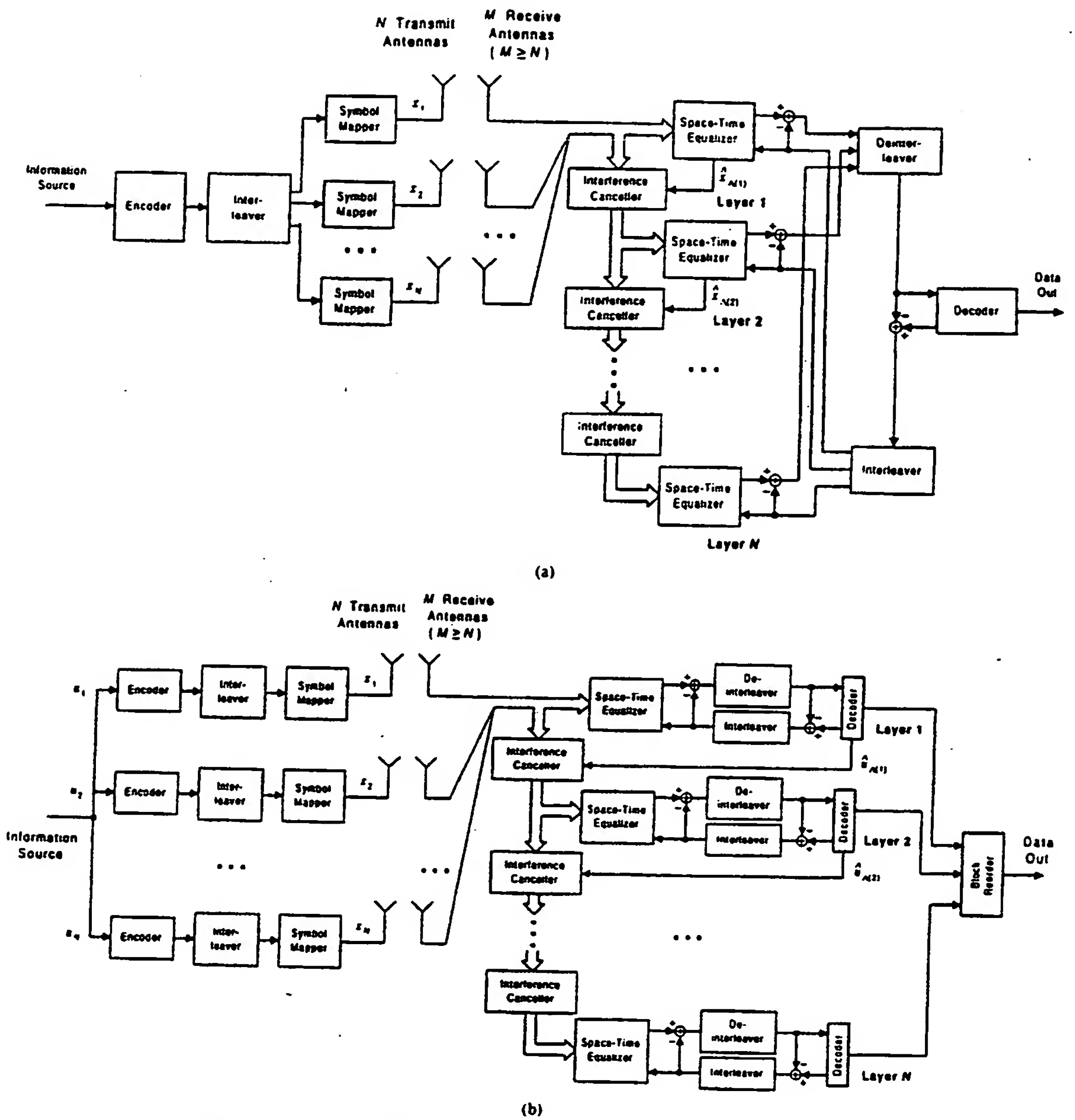


Fig. 1. Coded layered space–time architecture: (a) LST-I and (b) LST-II.

N data streams are decoupled and independently deinterleaved and decoded. The output of LST-II produces N information blocks at a rate of $1/N$ times the output rate of LST-I.

In Fig. 1(a) and (b), "space–time equalizer" refers to a combined array processing (for interference nulling) and equalization function. Instead of the ZF criterion, we assume that the optimization of the antenna/equalizer weights is based on a MMSE criterion, which in general provides better performance than a ZF approach. Foschini [7] has also indicated a potential performance benefit of using MMSE (or "maximum SNR") rather than ZF in a layered space–time architecture. Although we show M receive antennas in Fig. 1(a) and (b) ($M \geq N$ is the sufficient condition for nulling $N-1$ interference), we only consider $M = N$ in this study.

Similar to [9], the underlying assumption of our layered space–time architecture is that the receiver can order the detections of N data streams such that an undetected layer always has the strongest received SNR. In LST-I, the space–time equalizer in each layer must provide data decisions $\hat{x}_{\lambda(i)}$ (λ denotes the permutation due to layer ordering) to the interference canceller, since decoding cannot be done until all the layers are processed. In LST-II, the interference cancellation in each layer can use more reliable data decisions $\hat{u}_{\lambda(i)}$ provided by the decoder. Thus, LST-I is more prone to decision errors than

LST-II. In order to minimize the effects of decision errors, and also to improve the joint detection/decoding performance in general, we assume the use of turbo processing in our layered space–time architecture. As shown in Fig. 1(a) and (b), the space–time equalizers and the decoders provide *extrinsic* soft information to one another by subtracting the received soft information from the newly computed soft information. Details on MMSE space–time equalization and turbo processing will be provided in the Section IV.

B. Capacity Analysis

Without getting into the detail of all the processing functions, we first discuss the general differences between the two coded layered space–time approaches. In particular, we are most interested in the *capacity combining* aspects of the two approaches.

Let SNR_k denote the output SNR of the array processing in the kth layer. First, we note that, in LST-II, the capacity of each processing layer is bounded by the spectral efficiency R of the modulation and coding in each layer, e.g., $R = 1$ for 8-PSK with rate 1/3 coding. Thus, the total capacity of LST-II is given by (similar to (1)–(3), we write capacity without showing the frequency dependence)

$$C_{\mathrm{LST-II}} = \sum_{k=1}^{N} \min\{R, \log_2[1 + \mathrm{SNR}_k]\}. \tag{9}$$

Without layer ordering, it is most likely that the overall performance of LST-II will be largely influenced by the error probability of the first processing layer with a diversity order of only 1. In contrast, our simulation results in Section V will show that LST-II with layer ordering can actually achieve a diversity order of approximately N.

Since coding is performed across all the processing layers in LST-I, the achieved SNR in each layer will contribute to the overall layer processing performance. As Foschini [7] indicated, such a coding scheme should be able to achieve the capacity lower bound in (3). Here, we provide a generalized formula of Foschini's lower bound by removing the ZF constraint and instead using SNR_k as the generalized output SNR.

$$C_F = \sum_{k=1}^{N} \log_2[1 + \mathrm{SNR}_k]. \tag{10}$$

Reference [6] provides output SNR formulas for different types of optimum space–time processors. Here, it is of great interest to express the capacity lower bound using the best performance achievable. In the following equation, we represent SNR_k in (10) by the "matched filter" bound-the maximum achievable SNR by any space–time processing receiver:

$$C_{F,MF} = \left\langle \sum_{k=1}^{N} \log_2[1 + \Gamma_k(f)] \right\rangle \tag{11}$$

where $\Gamma_k(f)$ is the "matched filter" bound[2] given by equation (15) in Section IV-A (simply a rewriting of the result in [6]).

[2]Note that the "matched filter" bound usually refers to the integrated SNR ($\Gamma_k(f)$) over the signal bandwidth (e.g., [6]). However, in the capacity context, we assume the best possible way to exploit the SNR's in all frequency components

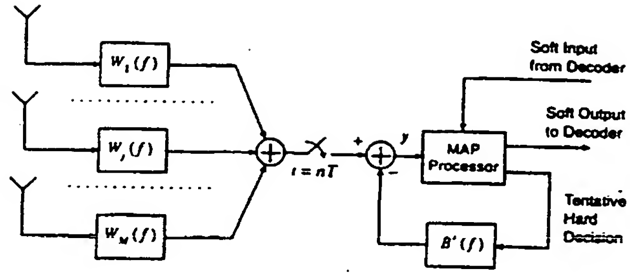


Fig. 2. Space-time DDFSE with MAP processing

Note that (11) is an explicit formula similar to (4); it shows the frequency dependence of the output SNR and the integration of capacity over the signal bandwidth. Also, we assume that the kth layer has $k - 1$ interferences.

In the process of analyzing the meaning of (11), we discovered an identical relationship between (11) and (4) regardless of how the layers are ordered. We show the proof in the Appendix (this proof is valid even when $M \neq N$). *Thus, Foschini's lower bound (3) is actually the true Shannon capacity bound when the output SNR of the space–time processing in each layer is represented by the corresponding "matched filter" bound.* This proves the optimality of the layered space–time concept.

The capacity analysis presented above is based on the assumption of perfect layer detection, i.e., no decision errors affecting the detection in subsequent layers. In reality, LST-I is more prone to decision errors than LST-II and layer ordering becomes important for both schemes. Our simulation results in Section V will demonstrate how decision errors affect the actual performance of the two coded layered space–time approaches.

IV. Signal Processing Functions

A. Space–Time Equalization

We consider combined array processing and equalization in order to cope with dispersive channels. A space–time equalizer, consisting of a spatial/temporal whitening filter, followed by a decision-feedback equalizer (DFE) or maximum-likelihood sequence estimator, can suppress both ISI and dispersive interference [6]. The space–time equalizer used in this study is shown in Fig. 2. It consists of a linear feedforward filter $W_j(f)$, $j = 1, \ldots, M$, on each diversity branch, a combiner, symbol-rate sampler, soft-input, soft-output (SISO) MAP sequence estimator, and synchronous linear feedback filter $B'(f)$. The feedforward filters $\{W_j(f)\}$ are shown as continuous-time filters, but they can be implemented in practice using fractionally-spaced tapped delay lines. The combined use of a sequence estimator and feedback filter after diversity combining is similar to the structure of a delayed decision-feedback sequence estimator (DDFSE) [27]. Thus, we refer to the space–time equalizer in Fig. 2 as a "space–time DDFSE." A "space–time DFE" is a structure where the sequence estimator is replaced by a memoryless hard slicer.

It has been shown in [20] that a space–time DDFSE with a sequence estimator memory of μ and a feedback filter of length $L_B - \mu$ can be optimized in a MMSE manner as if it was a space–time DFE with a feedback filter of length L_B. In fact,

numerical results in [6] showed that an optimum space–time DFE (with unconstrained filter lengths and no feedback decision errors) can perform within only 1–2 dB of the ideal "matched filter" bound performance. Thus, in order to have a practical receiver structure for layered space–time processing, we consider a space–time DDFSE with a minimum sequence estimator memory, i.e., $\mu = 1$. The sequence estimator is used only to provide a trellis structure needed for turbo processing, and presumably more reliable feedback decisions than the slicer used in a space–time DFE. Details on MAP processing will be given in Section IV-B.

We first provide a brief review of the space–time filtering theory. Based on the space–time DFE equivalent model described above, the MMSE solution for the feedforward filters $\{W_j(f)\}$ with unconstrained length can be given using the results of [6] (see also [28])

$$W = \Re_k^{-1} H_k^*(1 + B(f)) = \Re_{k-1}^{-1} H_k^* \frac{1 + B(f)}{1 + \Gamma_k(f)} \qquad (12)$$

where

$$\Re_k = \sum_{i=1}^{k} H_i^* H_i^T + \aleph \qquad (13)$$

$$W \triangleq \left[W_1\left(f - \frac{J}{T}\right) \cdots W_M\left(f - \frac{J}{T}\right) \cdots \right.$$
$$\left. \cdots W_1\left(f + \frac{J}{T}\right) \cdots W_M\left(f + \frac{J}{T}\right)\right]^T \qquad (14)$$

$$\Gamma_k(f) \triangleq H_k^T \Re_{k-1}^{-1} H_k^*. \qquad (15)$$

In the above equations, we assume that there are a total of k signal sources and we use H_k to indicate the channel vector [see (8)] of the *desired* signal. The remaining $k - 1$ signals are interferences. $B(f)$ in (12) is the feedback filter of the space–time DFE, which, from our assumption of $\mu = 1$, is only "1 tap" longer than $B'(f)$. $\Gamma_k(f)$ is the signal-to-interference-plus-noise power density ratio at frequency f, i.e., the "matched filter" bound. $B(f)$ can be determined through spectral factorization of $1 + \Gamma_k(f)$.

Equation (12) indicates that the optimum feedforward filter consists of a spatial/temporal filter $\Re_{k-1}^{-1} H_1^*$, which performs prewhitening ($\Re_{k-1}^{-1/2}$ is the whitening filter of interference and noise) and matching to the desired channel, followed by a temporal filter $(1 + B(f))/(1 + \Gamma_k(f))$, which is an anticausal post-whitening filter for suppressing precursor ISI.

A filter length analysis of the optimum space–time DFE described above is provided in [6]. We will first consider a finite-length realization of the space–time DFE based on the results presented there. We assume that the system has similar ISI situations as in EDGE and GSM. Namely, using the multipath delay profiles specified for EDGE and GSM (see Tables I and II), and assuming the same symbol rate of 270.833 kbaud ($T = 3.692$ μs) with Nyquist filtering (partial response signaling is used in EDGE and GSM), the ISI lasts up to five symbol periods for the hilly terrain (HT) profile in Table II. According to the empirical filter length formulas in [6], the feedforward filter on each branch should have the following causal and anticausal lengths to achieve near-optimum performance

$$L_C \approx K(N-1)(\rho_{\text{dB}}/10)$$
$$L_A \approx K + KN(\rho_{\text{dB}}/10) \qquad (16)$$

where K is the channel memory, N is used here to indicate the total number of signals, including the desired and interference signals, and ρ_{dB} is the average SNR in decibels. In our case, $K = 5$, and assume for example that the system has four transmit and four receive antennas ($N = 4$) and the operating range of average SNR is around 5 dB ($\rho_{\text{dB}} = 5$). The required filter length, including the *center* tap, will be $L_C + L_A + 1 \approx 23.5$. This is a highly impractical number, considering that four such filters are required, one per each receive antenna. Furthermore, as mentioned earlier, the optimum feedforward filters should be implemented using fractionally-spaced tapped delay lines. If a $T/2$-spaced filters are used, the total number of taps will be doubled. Such a space–time system with about 200 coefficients would be nearly impossible to compute in any radio link design.

Faced with such impracticality of an ideal signal processing arrangement, we proceed to consider a *suboptimum* option. First, we will use *symbol-spaced* instead of fractionally-spaced feedforward filters. In order to avoid significant performance penalties, a channel estimation-based timing recovery algorithm described in [29] will be used to optimize the symbol timing and the decision delay of the center tap relative to the measured channel impulse response. In principle, such timing optimization also allows the DFE to use a feedforward filter with a shorter span than the channel memory while achieving a reasonable performance [29], [30]. After experimenting with a number of significantly reduced filter length options, we decided on the following suboptimum space–time equalizer structure. The feedforward filter on each branch has a total of nine symbol-spaced taps, which are positioned such that $L_C = L_A = 4$. The feedback filter has a length of 8, i.e., $L_B = 9$ with the MAP processor memory $\mu = 1$ included (in order to completely cancel postcursor ISI, L_B must be at least as large as the channel memory plus the number of causal taps in the feedforward filter). The method in [29] is used to optimize the symbol timing and the decision delay of the center tap as described above. Direct matrix inversion is used to set

TABLE I
GSM TYPICAL URBAN (TU) CHANNEL MODEL

Path Delay (μs)	0.0	0.2	0.5	1.6	2.3	5.0
Path Power (dB)	−3.0	0.0	−2.0	−6.0	−8.0	−10.0

TABLE II
GSM HILLY TERRAIN (HT) CHANNEL MODEL

Path Delay (μs)	0.0	0.2	0.4	0.6	15.0	17.2
Path Power (dB)	0.0	−2.0	−4.0	−7.0	−6.0	−12.0

all the filter coefficients in a standard MMSE linear processing fashion [4], [26], [31], assuming perfect channel estimation.

B. Turbo Processing

The turbo processing technique used in this study is also based on a standard approach—the reader is referred to the rich literature [11]–[19], [21]–[22], [32]–[35] for a thorough treatment of this subject. The space–time equalizer and the decoder both performs SISO sequence estimation to compute the *a posteriori* probability (APP) of the transmit data symbols. This sequence estimation is done using the Bahl–Cocke–Jelinek–Raviv (BCJR) forward/backward algorithm. In the following, we describe the basic principle of the iterative detection/decoding process.

Using the BCJR algorithm, the MAP processor in the space–time equalizer with m^{μ} states (m is the signal constellation size, e.g., $m = 8$ for 8-PSK, and $\mu = 1$ in our case) computes the APP $P[c_k|y]$ of the kth coded bit c_k based on the observation y, where y is the equalizer output sequence corresponding to all the data symbols in a received block (see Fig. 2), and the *a priori* information provided by the decoder (this is not available in the first "turbo" iteration). The logarithm $\lambda(c_k) \triangleq \log_e(P[c_k|y])$ of this APP can be regarded as the sum of two terms

$$\lambda(c_k) = \lambda^p(c_k) + \lambda^e(c_k) \tag{17}$$

where $\lambda^p(c_k) \triangleq \log_e(P[c_k])$ is the logarithm of the *a prior* information provided by the decoder, and $\lambda^e(c_k)$ is called the "extrinsic" information. In each "turbo" iteration, the space–time equalizer subtracts $\lambda^p(c_k)$ from the newly computed value of $\lambda(c_k)$ to obtain the extrinsic information $\lambda^e(c_k)$ [see Fig. 1(a) and (b)]. The entire sequence $\{\lambda^e(c_k)\}$ is deinterleaved and forwarded to the decoder.

Similarly, the decoder computes the log-APP $v(c_k) \triangleq \log_e(P[c_k|\{\lambda^e(c_k)\}])$ based on the deinterleaved extrinsic information provided by the space–time equalizer, and subtract $\lambda^e(c_k)$ from it to obtain extrinsic information $v^e(c_k)$. The extrinsic information is then interleaved and forwarded to the equalizer as the new *a priori* information $\lambda^p(c_k)$ for the next "turbo" iteration.

The interleaver considered in this study is a *pseudo-random* interleaver, i.e., we generate a pseudo-random permutation of numbers from 1 to l, where l is the block length, and then use this permutation as a *fixed* interleaver.

In combining the branch metric obtained from the equalizer output with the branch metric obtained from the soft input provided by the decoder, the MAP processor in the space–time equalizer must compute the *a priori* information for each 8-PSK symbol x_k from the three soft inputs ($\lambda^p(c_{3k})$, $\lambda^p(c_{3k+1})$, $\lambda^p(c_{3k+2})$). We assume that this is done by way of summing the three soft inputs as if the three coded bits were transmitted from independent sources (these soft inputs are actually not independent when conditioned on the observed waveform of the entire data burst). This is a suboptimal method, which is known to cause a "random modulation" performance degradation effect in bit-interleaved coded modulation [36]–[38]. However, this effect can be overcome by iterative decoding [38], which is implicit in our turbo space–time processing approach.

As noted earlier, in LST-I, the space–time equalizer in each layer must provide immediate data decisions to be used for interference cancellation. Since these decisions are not "protected" by coding, they are prone to errors. In this study, we explore a soft decision technique to minimize the effect of decision errors. The optimum soft decision can be computed by averaging all the possible transmit symbols weighted by their APP's [39]

$$\hat{x}_k = \sum_{x \in X} sP[x_k = x|y] \tag{18}$$

where X includes all the complex-valued 8-PSK constellation points. Since $P[x_k = x|y]$ can be obtained along with the computation of the APP $P[c_k|y]$, this soft decision approach can be implemented with nearly no additional cost in complexity. Similarly, we apply the same technique to compute soft decision outputs in LST-II.

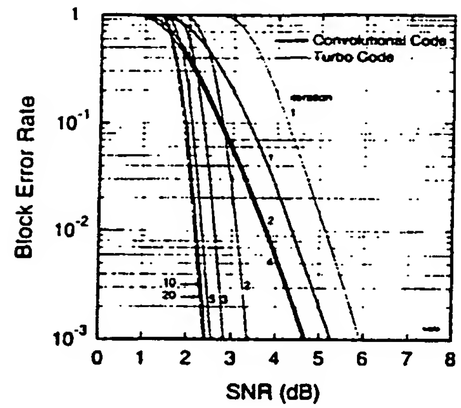
V. Performance Results

A. Performance Criteria and System Assumptions

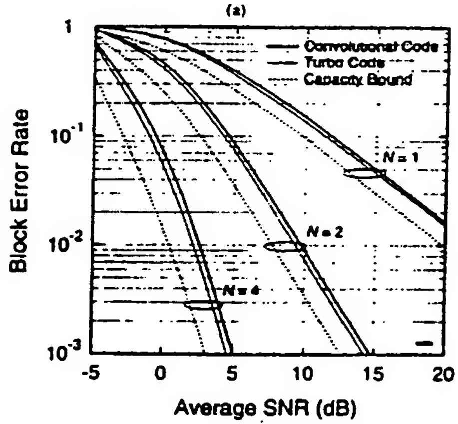
We now present performance results of the layered space–time concepts described so far. The performance measure is the block-error rate (BLER) over Rayleigh fading. The results are obtained through Monte Carlo simulation. The BLER is averaged over up to 40 000 channel realizations. Each block contains 400 information bits (before coding).

In comparing the performance results to channel capacity, we follow the convention of a number of previous works (e.g., [9], [23]) to compare the computed BLER with the "outage capacity" [1], i.e., the probability that a specified bit rate is not supported by the channel capacity. This is a vague comparison, since the Shannon limit refers to the highest error-free bit rate possible for long encoded blocks but it does not specify how long the blocks should be. Nevertheless, such a comparison should still be meaningful as long as the block length and BLER are specified. This is similar to the way a bit-error rate of 10^{-5} is commonly used as the "error free" reference for an additive white Gaussian noise (AWGN) channel.

In order to assess the best performance achievable, we assume that the channel characteristics can be perfectly estimated at the receiver. Similarly, the choices of 1-D processing and coding techniques are important to deliver the best possible performance. We try to optimize these choices while keeping them as practical as possible. Except for the use of array processing and iterative MAP algorithms, all the radio link techniques assumed in this study are "legacy-compatible" with the EDGE standard (note also that vast research interest in turbo coding has made simplified MAP algorithms available [21], [22] that are not much more complex than the conventional Viterbi algorithm). None of these techniques are claimed to be optimum. Yet, our results indicate that near-capacity performance is achievable when combining them through the coded layered space–time architectures.



(a)



(b)

Fig. 3. Performance of bit-interleaved 8-PSK with rate-1/3 convolutional and turbo coding over (a) AWGN channel, and (b) quasi-static flat Rayleigh fading channels with N receive diversity antennas.

B. Choosing the Coding Scheme

We consider a bit-interleaved coded modulation scheme using 8-PSK with Gray mapping and rate-1/3 coding. Square-root Nyquist filtering with 30% rolloff is assumed at the transmitter and receiver. Bit-interleaved coded modulation has been shown [36], [37] to outperform traditional trellis-coded modulation in fast fading channels (where *time* diversity can be exploited through sufficient interleaving) and it can be improved upon by considering a better mapping technique that permits a large Euclidean distance without sacrificing the maximum Hamming distance of the baseline coding scheme [38]. In this paper, though, since quasi-static fading is our basic assumption, the code by itself must be able to withstand deep fades. In principle, any code that performs well in an AWGN channel is considered a good candidate—turbo codes are among the strongest candidates that come to mind.

Fig. 3(a) and (b) provide a performance comparison between two rate-1/3 coding schemes: one using a 64-state convolutional code with (octal) generators $(G_1, G_2, G_3) = (155, 117, 123)$ (the same code as proposed for EDGE [20]) and the other using a turbo code with two identical 16-state recursive encoders similar to the scheme originally proposed by Berrou and Glavieux [12] (the results here assume generators $(G_1, G_2) = (23, 31)$, which appeared to perform slightly better than other generators we tested). Both schemes assume the use of bit-interleaved 8-PSK with Gray mapping. The turbo coding scheme has an additional interleaver within the encoder, which uses another pseudo-randomly generated permutation. The receiver structure is consistent with what we have described so far. Note, however, that we assume a minimum number of filter taps (only one feed-forward tap per branch and no feedback filter) whenever there is no delay spread assumed, although the MAP processor in the DDFSE is always used for iterative detection/decoding as described in Section IV-B. For the turbo coding scheme, "one iteration" means a full cycle of three processes: 1) MAP processing in DDFSE; 2) turbo decoding by the first decoder; and 3) turbo decoding by the second decoder.

Fig. 3(a) shows the performance of the two coding schemes in an AWGN channel. First, we note that the performance of convolutional coding also benefits from iterative processing. This is due to the suboptimal nature of the decoding scheme, i.e., the "random modulation" effect described earlier, which can be improved through iterative decoding. Fig. 3(a) shows that most of the improvement is achieved within two decoding cycles. For turbo coding, the performance still improves even after five iterations, but saturates quickly after ten iterations. At 10^{-3} BLER (approximately equivalent to 10^{-5} bit-error rate), turbo coding outperforms convolutional coding by about 2.2 dB, and the required SNR is within only 2.4 dB of the 0-dB Shannon limit for a spectral efficiency of 1 bps/Hz (8-PSK with rate-1/3 coding).

However, when we look at the average BLER performance over quasi-static flat fading channels in Fig. 3(b), the benefit of turbo coding (with ten iterations) over convolutional coding (with two iterations) is reduced to only about 0.5 dB at any value of the average SNR and for all the assumed numbers of receive diversity antennas. This is not surprising for two reasons: First, it is well known that the average BLER is determined mostly by the probability of fading events that results in high BLER's. If we look at the relative performance at a BLER of, say, above 10% in Fig. 3(a), the difference between the two coding schemes is indeed less than 1 dB. Second, the performance of convolutional coding over fading channels is already within about 2 dB of the capacity bound—the capacity bound in this case is defined as the probability that the combined output SNR of all diversity branches is below the 0-dB Shannon limit. Thus, there is not much room for further improvement.

Based on the fact that the performances of the two coding schemes are quite similar in quasi-static fading channels, we will only consider convolutional coding in the remainder of this paper.

C. Layered Space–Time Performance

We first look at the performance of the two coded layered space–time approaches over a flat Rayleigh fading channel. Fig. 4 shows the different capacity bounds for this channel, assuming $N = 2, 4,$ and 8, where N is the number of transmit and receive antennas. Again, although we plot the results as "block error rate," the capacity bound is defined as the probability that the specified spectral efficiency R ($R = N$ in this case) is not supported by each of the differently defined channel capacities. C denotes the Shannon capacity bound given by (4)

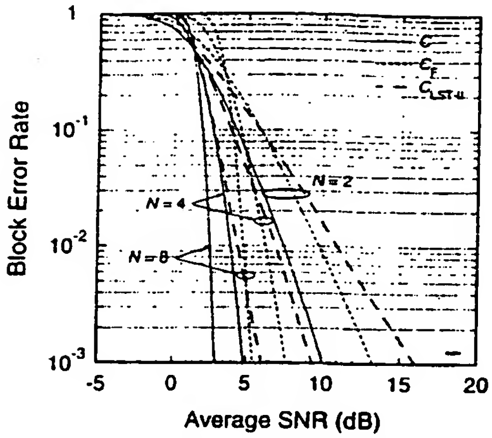


Fig. 4. Capacity bounds for quasi-static flat Rayleigh fading channels with N transmit and N receive antennas. C: Shannon capacity, C_F: Foschini (original) bound, and C_{LST-II}: capacity bound for LST-II.

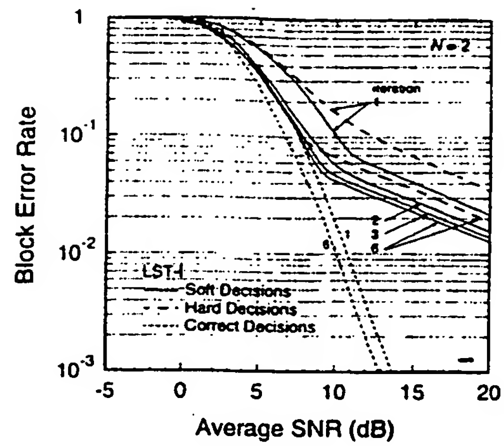


Fig. 5. Layered space–time performance of LST-I with two transmit and two receive antennas ($N = 2$). Quasi-static flat Rayleigh fading channel.

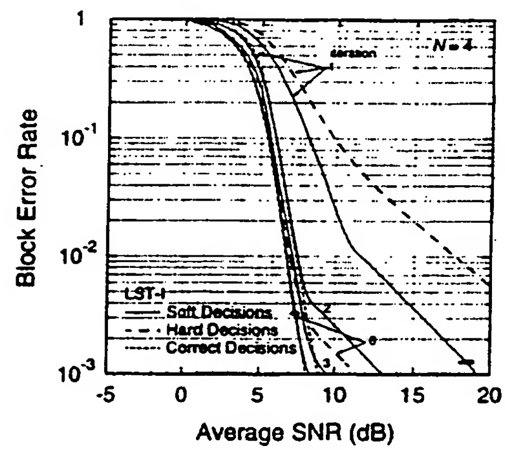


Fig. 6. Layered space–time performance of LST-I with four transmit and four receive antennas ($N = 4$). Quasi-static flat Rayleigh fading channel.

[which is equivalent to the generalized Foschini bound $C_{F,MF}$ in (11)]. C_F denotes the original Foschini bound (with the ZF constraint) in (3), and C_{LST-II} is the capacity bound for LST-II in (9) (however, the results for C_{LST-II} are obtained simply by averaging the probability that $R = 1$ is not supported by each processing layer). Note that C_{LST-II} can indeed provide approximately a diversity order of N; this is attributed largely to the use of layer ordering as discussed earlier. Note also that all the bounds show an improvement with increasing N. This means that the capacity actually increases more than linearly with the number of transmit and receive antennas. However, there is a diminishing improvement as N increases to a much larger number.

Fig. 5 shows the simulation results for LST-I with 2 transmit and 2 receive antennas (i.e., $N = 2$). Three sets of results are provided, assuming: 1) soft decisions; 2) hard decisions; and 3) correct decisions in each layer (note that the DDFSE always uses tentative decisions and provides soft outputs to the decoder). We see that, although soft decisions offer some improvement over hard decisions, the impact of decision errors is still quite noticeable. Fig. 6 shows similar results for $N = 4$. Here, the impact of decision errors is not as significant as the previous results, and turbo processing and soft decisions help to reduce much of this impact. With three iterations, the effect of decision errors almost completely disappears when using soft decisions. Decision errors have a lesser effect for a larger N because of the greater diversity order available through array processing and layer ordering.

In Fig. 7, we compare the results using soft decisions and six "turbo" iterations with the Shannon capacity bound. For $N = 4$ and 8, the performance of LST-I is within 2.5–3 dB of the capacity bound at 10% BLER (and about 3–3.5 dB at 1% BLER). Since the BLER may vary as a function of the block size,[3] it is also important to consider the *processing* loss by discounting the loss due to the inefficiency of modulation and coding. As shown in Fig. 3(b), there is already a gap of about 2 dB between the performance of our coding scheme and the Shannon limit.

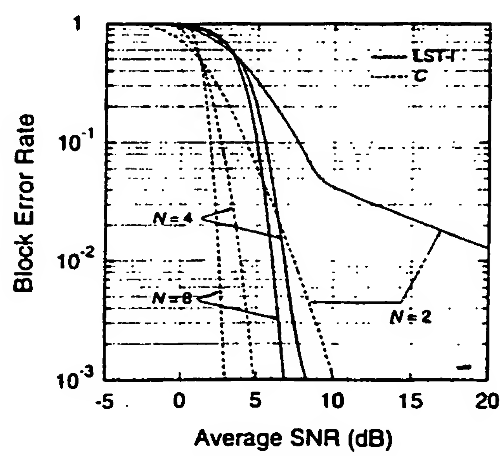


Fig. 7. Layered space–time performance of LST-I for $N = 2$, 4, and 8. Soft decisions. 6 iterations. Quasi-static flat Rayleigh fading channel.

[3]As an example, when we double the block size, the required average SNR is 0.2–0.4 dB greater than the results shown here. However, this difference in average SNR applies uniformly to all results, with or without layered space–time processing.

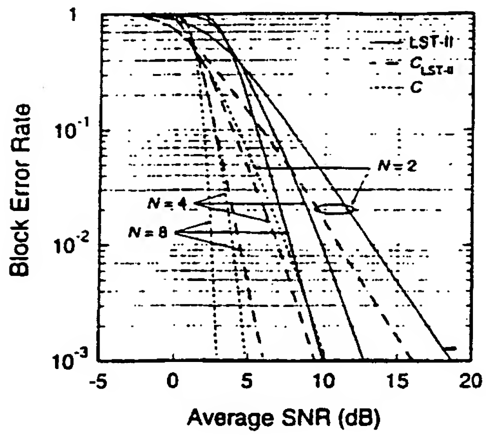


Fig. 8. Layered space–time performance of LST-II for $N = 2$, 4, and 8. Soft decisions. 2 iterations. Quasi-static flat Rayleigh fading channel.

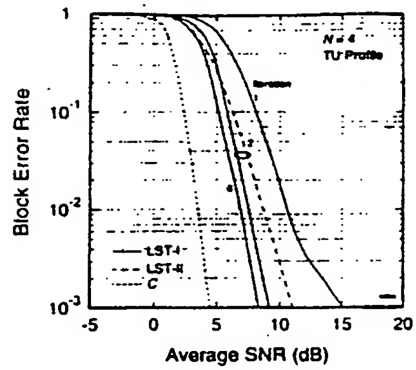
Thus, the actual loss due to layered space–time processing is only about 0.5–1 dB at 10% BLER.

Fig. 8 shows the performance of LST-II for $N = 2$, 4, and 8, compared to the Shannon and C_{LST-II} bounds. Soft decisions and two "turbo" iterations are assumed. (In this case, we found the effect of decision errors to be marginal, i.e., the results with hard and correct decisions were generally within 1 dB of the results shown here. Also, we found turbo processing with more than two iterations to provide little improvement.) At 10% BLER, the performance of LST-II is 2.5 dB from the C_{LST-II} bound, and 3.5 dB from the Shannon bound. At 1% BLER, however, the loss compared to the Shannon bound can be as much as 6 dB.

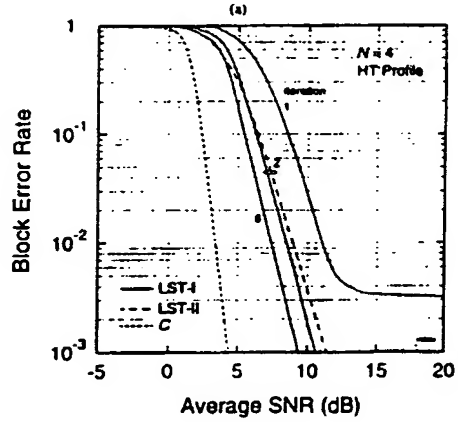
From the above results, we conclude that, for $N = 4$ and 8, LST-I outperforms LST-II by a margin of 0.5 dB (at 10% BLER) to 3 dB (at 1% BLER). For $N = 2$, the performance of LST-I is greatly affected by decision errors (note that, even in this case, LST-I still performs as well as LST-II at 10% BLER), whereas LST-II can reach a lower BLER at high average SNR. Based on these results, the layered space–time approach is not highly recommended for $N = 2$. As mentioned earlier, space–time coding is a better alternative to achieve a spectral efficiency of 2 bps/Hz. For instance, a 64-state space–time coded QPSK can perform to within 2 dB of the Shannon capacity bound [23].

D. Frequency-Selective Channels

Finally, we present an example of performance results for frequency-selective fading channels. This example assumes $N = 4$ and the use of soft decisions for both LST-I and LST-II. Fig. 9(a) and (b) show the results for the TU and HT profiles, defined in Tables I and II. Again, we only show results with two "turbo" iterations for LST-II because little improvement can be achieved with more iterations in this case. For both delay profiles, the performance at 10% BLER is within 3 dB of the Shannon bound for LST-I with six iterations, and within 4 dB for LST-II with two iterations. At a lower BLER, the loss relative to the bound is greater for HT than for TU. This is due to the limitation of



(a)



(b)

Fig. 9. Layered space–time performance of LST-I and LST-II over frequency-selective channels: (a) TU profile, (b) HT profile. $N = 4$. Soft decisions.

the suboptimum space–time equalizer structure we assume, as already discussed in Section IV-A.

VI. Conclusion

By deriving the generalized Shannon capacity formula and suggesting a layered space–time architecture that attains a tight lower bound on the capacity achievable, Foschini has laid a significant theoretical foundation for improving the wireless channel capacity through multiple-element array technology. We have shown that Foschini's lower bound is actually the true Shannon bound when the output SNR of the space–time processing in each layer is represented by the corresponding "matched filter" bound. We then provided two coded layered space–time approaches as an embodiment of this concept. For a large number of transmit and receive antennas, coding across the layers provides a better performance than independent coding within each layer. However, with two transmit and two receive antennas, the former is heavily affected by decision errors and, therefore, provides a poorer performance than the latter.

The underlying coding and signal processing techniques used in this study are based on practical but suboptimal approaches. Yet, such suboptimality can be greatly compensated for by it-

erative processing. Overall, our coded layered space–time approaches can achieve a performance within about 3 dB of the Shannon bound at 10% BLER, about 2 dB of which is a loss due to the practical coding scheme we assume. Thus, not only is the layered space–time architecture exactly what the Shannon limit has prescribed in a theoretical sense, but it also provides an attractive general methodology for improving and achieving the wireless channel capacity.

APPENDIX

PROVING THE EQUIVALENCE OF FOSCHINI BOUND AND SHANNON CAPACITY

Using the mathematical induction method, we will prove that (4) and (11) are identical. In order to do so, we must show that

$$\det(\Re\aleph^{-1}) = \prod_{k=1}^{N}(1+\Gamma_k(f)) \tag{19}$$

where $\Re$ in the above equation is equivalent to $\Re_N$ defined in (13). Again, we assume that the kth layer has $k-1$ interferences. Note also that the proof provided here is independent of the number of receive antennas M (the dimension of $\Re$) and the way the layers are ordered.

We start by assuming 1 signal source and M receive antennas. It can be easily shown [1] that

$$\det(\Re_1\aleph^{-1}) = 1 + H_1^T\aleph^{-1}H_1^* = 1+\Gamma_1(f). \tag{20}$$

Next, we assume that (19) is true for the case of $n-1$ signal sources, i.e.,

$$\det(\Re_{n-1}\aleph^{-1}) = \prod_{k=1}^{n-1}(1+\Gamma_k(f)). \tag{21}$$

We then show in the following that, given (21), (19) is also true for n signal sources.

First, we note from (13) that

$$\Re_n = \Re_{n-1} + H_n^* H_n^T. \tag{22}$$

Using the matrix inversion lemma [4, Appendix D], we can show that [similar to (12)]

$$\Re_n^{-1}H_n^* = \frac{\Re_{n-1}^{-1}H_n^*}{1+\Gamma_n(f)}. \tag{23}$$

It follows that

$$(\Re_n\aleph^{-1})H_n^* = \aleph\Re_n^{-1}H_n^* = \frac{\aleph\Re_{n-1}^{-1}H_n^*}{1+\Gamma_n(f)} = \frac{(\Re_{n-1}\aleph^{-1})^{-1}H_n^*}{1+\Gamma_n(f)}. \tag{24}$$

For convenience, let

$$A \triangleq \Re_n\aleph^{-1} \text{ and } B \triangleq \Re_{n-1}\aleph^{-1}. \tag{25}$$

We can rewrite (24) as

$$A^{-1}H_n^* = \frac{B^{-1}H_n^*}{1+\Gamma_n(f)}. \tag{26}$$

Furthermore, using the matrix identity [40]

$$A^{-1} = \frac{A_{\mathrm{adj}}}{\det(A)} \tag{27}$$

where A_{adj} is called the *adjugate matrix* of matrix A, we can rewrite (26) as

$$\frac{A_{\mathrm{adj}}}{\det(A)}H_n^* = \frac{B_{\mathrm{adj}}}{\det(B)}\,\frac{H_n^*}{1+\Gamma_n(f)}. \tag{28}$$

By replacing $\det(B)$ in the above equation using (21), we obtain

$$\frac{A_{\mathrm{adj}}H_n^*}{\det(A)} = \frac{B_{\mathrm{adj}}H_n^*}{\prod_{k=1}^{n}(1+\Gamma_k(f))}. \tag{29}$$

Our goal is to prove that

$$\det(A) = \prod_{k=1}^{n}(1+\Gamma_k(f)). \tag{30}$$

Thus, given (29), we must show that

$$A_{\mathrm{adj}}H_n^* = B_{\mathrm{adj}}H_n^*. \tag{31}$$

From (22) and (25), we have

$$\begin{aligned} A &= B + H_n^* H_n^T\aleph^{-1} \\ &= \left[b_1 + H_n^*\frac{H_{n1}(f)}{\aleph_1(f)},\, b_2 + H_n^*\frac{H_{n2}(f)}{\aleph_2(f)},\, \cdots,\right. \\ &\qquad \left. b_M + H_n^*\frac{H_{nM}(f)}{\aleph_M(f)}\right] \end{aligned} \tag{32}$$

where b_j is the jth column vector of B, for $j = 1, \cdots, M$; for convenience, we use M here to indicate the *overall* receive diversity order, including the effects of both multiple antennas and excess bandwidth.

We now prove (31) by showing that the jth element of $A_{\mathrm{adj}}H_n^*$ is equal to the jth element of $B_{\mathrm{adj}}H_n^*$, for $j = 1, \ldots, M$. Note that the jth element of $A_{\mathrm{adj}}H_n^*$ is given by $\det(\overline{A}_j)$, where $\overline{A}_j$ is obtained by replacing the jth column of A by H_n^*. Similarly, the jth element of $B_{\mathrm{adj}}H_n^*$ is given by $\det(\overline{B}_j)$, where $\overline{B}_j$ is obtained by replacing the jth column of B by H_n^*.

Using (32) and the linear properties of determinants, we can show that

$$\det(\overline{A}_j) = \det\left[b_1 + H_n^*\frac{H_{n1}(f)}{\aleph_1(f)},\, \cdots,\, \underset{H_n^*}{\overset{j\text{th column}}{}},\, \cdots\cdots\right] \tag{33}$$

$$\begin{aligned} &= \det[b_1, \cdots, H_n^*, \cdots\cdots] \\ &\quad + \det\left[H_n^*\frac{H_{n1}(f)}{\aleph_1(f)}, \cdots, H_n^*, \cdots\cdots\right] \\ &= \det[b_1, \cdots, H_n^*, \cdots\cdots] \\ &\quad + \frac{H_{n1}(f)}{\aleph_1(f)}\det[H_n^*, \cdots, H_n^*, \cdots\cdots] \\ &= \det[b_1, \cdots, H_n^*, \cdots\cdots]. \end{aligned} \tag{34}$$

Similarly, we can expand the result of (34) with respect to the second column, the third column, and so on (except for the jth column). Eventually, we obtain

$$\det(\overline{A}_j) = \det[b_1, b_2, \cdots, \overset{j\text{th column}}{H_n^*}, \cdots, b_M] = \det(\overline{B}_j) \tag{35}$$

which proves (31). The proof of (19) is therefore complete.

ACKNOWLEDGMENT

The author wishes to thank G. J. Foschini for reviewing the information theory part of the original manuscript of this paper and suggesting several improvements, and for the enormous inspiration he provided by inventing the layered space–time concept. The author also benefited from discussions with K. R. Narayanan, I. Lee, and X. Li. Finally, the author thanks the anonymous reviewers for valuable comments and suggestions.

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According to the present invention, the receiver can select a set of antenna elements, including their number and / or identity, from among a larger group of antenna elements in order, among other things, to improve performance of the system without increasing the extent of radiofrequency circuitry. One process for selecting antenna elements is to utilize equation 4 above as a measure of quality for the particular set of antenna elements being evaluated. That evaluation can occur for each permutation or combination of antenna elements in order to select the subset with optimum performance (as determined, for instance, by selecting the subset with greatest value calculated according to equation 4). This can occur at whatever desired points in time, including periodically.

CLAIMS

What is claimed is:

1. A radio receiver coupled to a plurality of receiver antenna elements, comprising a plurality of layers for processing signals received by the receiver elements, each layer comprising:

a space-time equalizer, the space-time equalizer in a first layer of the plurality of layers coupled to each of the receiver elements, the space-time equalizer in each of the other layers of the plurality of layers coupled to an interference canceller which receives output from the layer preceding each said other layer.

2. A receiver according to claim 1 further comprising a deinterleaver coupled to each space time equalizer and to a decoder for output.

3. A radio receiver according to claim 1 in which each layer comprises its own deinterleaver and decoder, and further comprises an interleaver adapted to receive output from the decoder in said layer and from the deinterleaver in said layer, the interleaver feeding output to the equalizer in said layer, thereby allowing soft decisions about information being processed iteratively by said layer to be fed back and forth between said equalizer and said decoder in said layer.

4. A radio receiver according to claim 3 in which the interference canceller which receives output from the layer preceding each said other layer receives information from the decoder and the interference canceller in the preceding layer.

5. A radio receiver according to claim 1 in which the equalizer in each layer is connected to a common deinterleaver, which feeds a common

decoder, and in which a common interleaver receives signals from the decoder, and is coupled to each equalizer in order to provide interleaved signals to the equalizer.

6. A radio receiver according to claim 5 in which the interference canceller which receives output from the layer preceding each said other layer receives signals from the equalizer and from the interference canceller in said layer preceding each said other layer.

7. A radio receiver according to claim 1 in which each space-time equalizer performs equalization using a minimum mean-square error criterion.

8. A radio receiver according to claim 1 in which the receiver is adapted to receive and process signals from a transmitter that is coupled a plurality of transmit antenna elements, the number of transmit antenna elements to which the transmitter is coupled being less than the number of receiver antenna elements to which the receiver is coupled.

9. A radio receiver according to claim 8 in which the receiver is adapted to receive and process signals from a transmitter that is coupled to N transmit antenna elements, the number of receiver antenna elements to which the receiver is coupled is M, and M is greater than N.

10. A radio receiver according to claim 9 adapted to be coupled to at least one set of M receiver antenna elements out of K available receiver antenna elements.

11. A radio receiver according to claim 1 in which the receiver is adapted to select the sequence in which information from the receiver antenna elements is to be processed.

12. A communications system, comprising:

a radio transmitter coupled to a stream of information and to a plurality of transmit antenna elements, said transmitter adapted to apportion a portion of the stream of information to each transmit antenna element by interleaving said portions of the information stream among the transmit antenna elements; and

a radio receiver coupled to a plurality of receiver antenna elements, comprising a plurality of layers for processing signals received by the receiver elements, each layer comprising:

a space-time equalizer, the space-time equalizer in a first layer of the plurality of layers coupled to each of the receiver elements, the space-time equalizer in each of the other layers of the plurality of layers coupled to an interference canceller which receives output from the layer preceding each said other layer.

13. A system according to claim 12 further comprising a deinterleaver coupled to each space time equalizer and to a decoder for output.

14. A system according to claim 12 in which each layer comprises its own deinterleaver and decoder, and further comprises an interleaver adapted to receive output from the decoder in said layer and from the deinterleaver in said layer, the interleaver feeding output to the equalizer in said layer, thereby allowing soft decisions about information being processed iteratively by said layer to be fed back and forth between said equalizer and said decoder in said layer.

15. A system according to claim 14 in which the interference canceller which receives output from the layer preceding each said other layer receives information from the decoder and the interference canceller in the preceding layer.

16. A system according to claim 12 in which the equalizer in each layer is connected to a common deinterleaver, which feeds a common decoder, and in which a common interleaver receives signals from the deinterleaver and decoder, and is coupled to each equalizer in order to provide interleaved signals to the equalizer.

17. A system according to claim 16 in which the interference canceller which receives output from the layer preceding each said other layer receives signals from the equalizer and from the interference canceller in said layer preceding each said other layer.

18. A system according to claim 12 in which each space-time equalizer performs equalization using a minimum mean-square error criterion.

19. A system according to claim 12 in which the number of transmit antenna elements to which the transmitter is coupled is less than the number of receiver antenna elements to which the receiver is coupled.

20. A system according to claim 12 in which the transmitter is coupled to N transmit antenna elements, the receiver is coupled to M antenna elements, and M is greater than N.

21. A system according to claim 20 in which the receiver is adapted to be coupled to at least one set of M receiver antenna elements out of K available receiver antenna elements.

22. A system according to claim 12 in which the receiver is adapted to select the sequence in which information from the receiver antenna elements is to be processed.

23. A communications system, comprising:

a radio transmitter coupled to a stream of information and to a plurality of transmit antenna elements, said transmitter adapted to apportion a portion of the stream of information to each transmit antenna element by interleaving said portions of the information stream among the transmit antenna elements; and

a radio receiver coupled to a plurality of receiver antenna elements, comprising a plurality of layers for processing signals received by the receiver elements, each layer comprising:

a space-time equalizer, a deinterleaver and a decoder, the space-time equalizer in a first layer of the plurality of layers coupled to each of the receiver elements, the space-time equalizer in each of the other layers of the plurality of layers coupled to each of the receiver elements and to an interference canceller which receives output from the decoder in the layer preceding each said other layer, each space-time equalizer coupled to the deinterleaver in its layer, each deinterleaver coupled to the decoder in its layer;

the equalizer in each layer adapted to perform minimum mean-square error equalization to signals being processed; and

an output for said stream of information coupled to the decoders in each of said layers.

24. A system according to claim 23 in which the transmitter is adapted to interleave portions of the information stream among the transmit antenna elements randomly.

25. A system according to claim 23 in which the transmitter is adapted to interleave portions of the information stream among the transmit antenna elements pseudo-randomly.

26. A system according to claim 23 in which the number of transmit antenna elements to which the transmitter is coupled is less than the number of receiver antenna elements to which the receiver is coupled.

27. A system according to claim 26 in which the transmitter is coupled to N transmit antenna elements, the receiver is coupled to M antenna elements, and M is greater than N.

28. A system according to claim 27 in which the receiver is adapted to be coupled to at least one set of M receiver antenna elements out of K available receiver antenna elements.

29. A system according to claim 23 in which the receiver is adapted to select the sequence in which information from the receiver antenna elements is to be processed.

30. A process for communicating an information stream using a radio transmitter and a radio receiver, including:

a. coupling a radio transmitter to the information stream and to a plurality of transmit antenna elements, and interleaving portions of the information stream among the transmit antenna elements;

b. transmitting said portions of the information stream;

c. coupling a radio receiver to a plurality of receiver antenna elements to receive the transmitted information stream; and

d. processing the information stream in a plurality of processing layers, comprising:

(i) in a first layer, coupling the receiver antenna elements to an equalizer and space-time processing the signals from the receiver antenna elements in the equalizer;

deinterleaving the output from the equalizer;

decoding the deinterleaved output from the equalizer;

feeding the decoded information to a common output for the information stream; and

(ii) in each successive layer, coupling to an equalizer the output from an interference canceller that is fed by output from the preceding layer and space-time equalizing said output from said interference canceller;

deinterleaving the output from the equalizer;

decoding the deinterleaved output from the equalizer;

and feeding the decoded information to a common output for the information stream.

31. A process according to claim 30 in which steps of deinterleaving the output from the equalizer, decoding the deinterleaved output from the equalizer and feeding the decoded information to a common output for the information stream are performed inside each layer and further comprising, in each layer, reinterleaving deinterleaved and decoded output from said layer, and feeding the reinterleaved signal to the equalizer in said layer thereby allowing soft decisions about information being processed iteratively by said layer to be fed back and forth between said equalizer and said decoder in said layer.

32. A process according to claim 30 in which the equalizer in each layer performs minimum mean-square error equalization.

33. A process according to claim 30 in which said interleaving is performed randomly.

34. A process according to claim 30 in which said interleaving is performed pseudo-randomly.

35. A process according to claim 30 in which coupling of said receiver to said receiver antenna elements includes selecting a set of receiver antenna elements from a larger group of receiver antenna elements.

36. A process according to claim 30 in which coupling of said receiver to said receiver antenna elements includes coupling to a set of receiver antenna elements selected from a larger group of receiver antenna elements.

37. A process according to claim 36 in which said coupling further includes selecting the sequence in which signals from receiver antenna elements are to be processed.

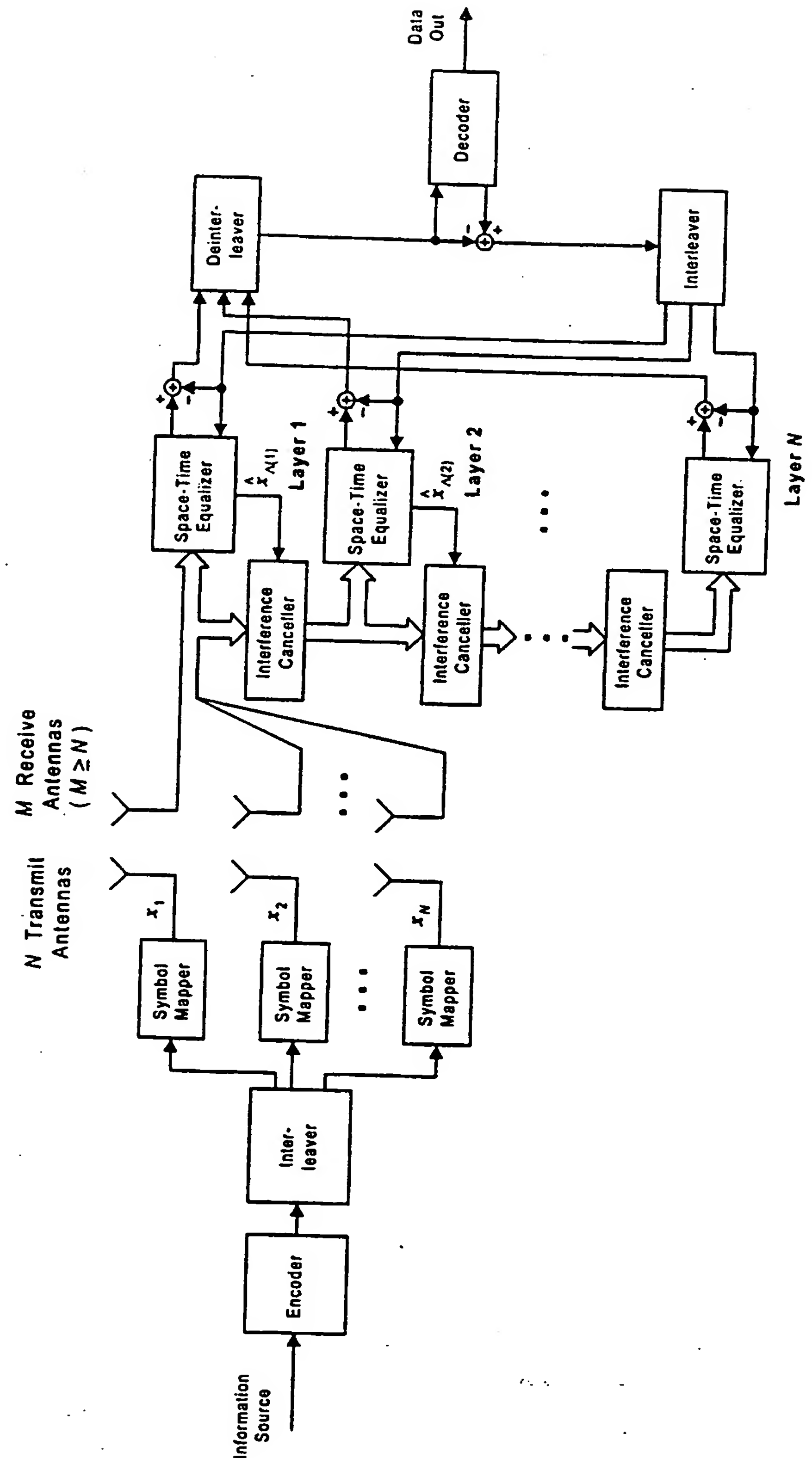


Fig. 1 (a) Coded layered space-time architecture: LST-I.

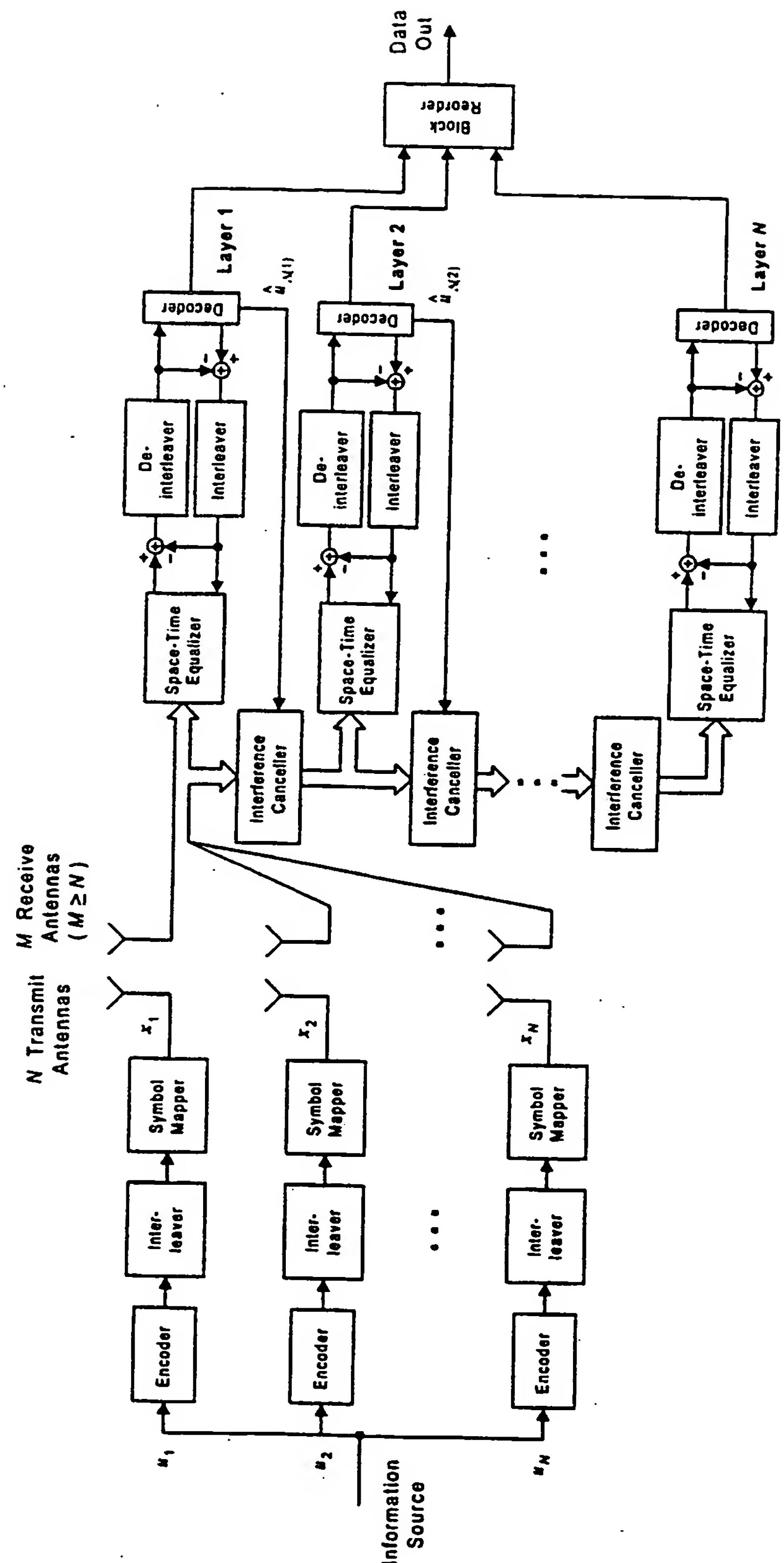


Fig. 1 (b) Coded layered space-time architecture: LST-II.

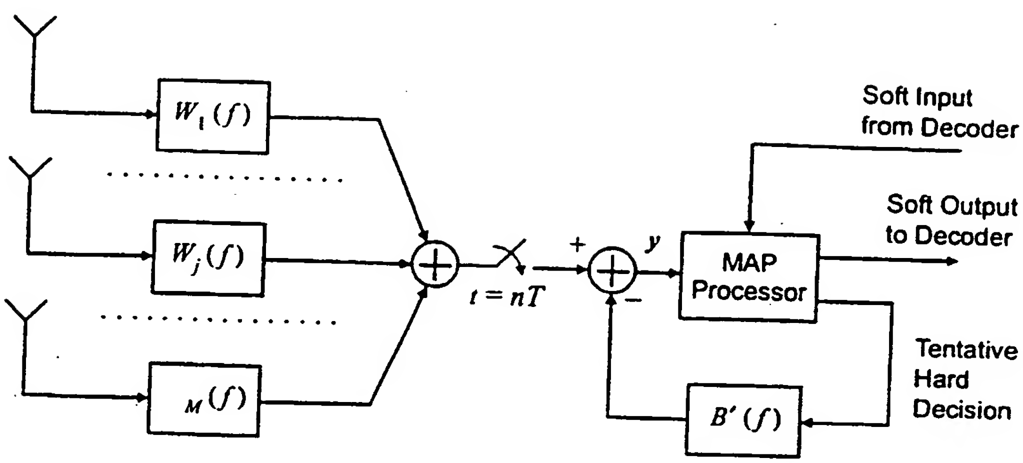
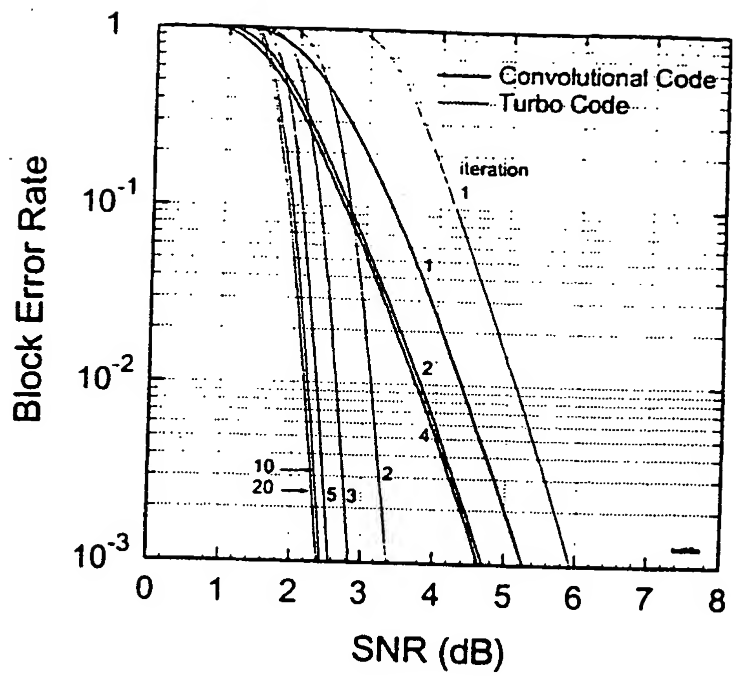
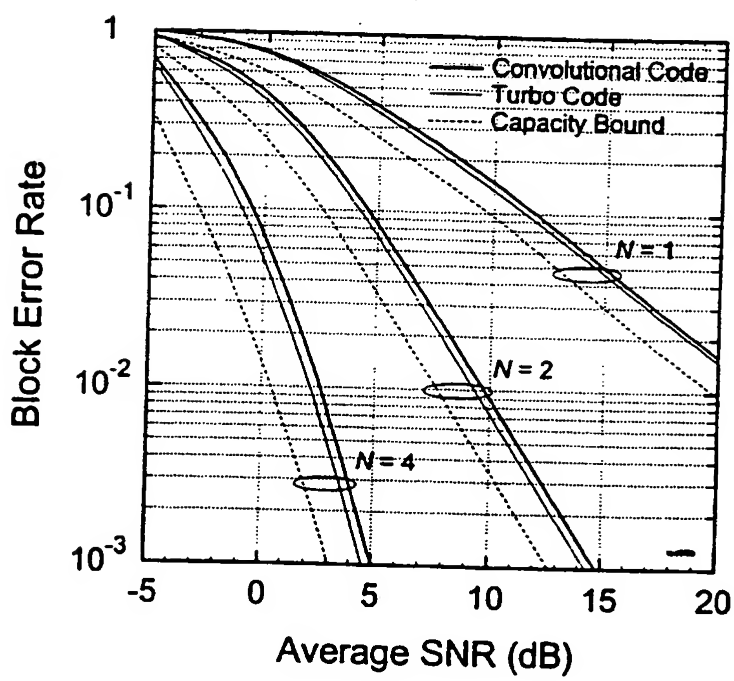


Fig. 2. Space-time DDFSE with MAP processing



(a)



(b)

Fig. 3. Performance of bit-interleaved 8PSK with rate-1/3 convolutional and turbo coding over (a) AWGN channel, and (b) quasi-static flat Rayleigh fading channels with N receive diversity antennas.

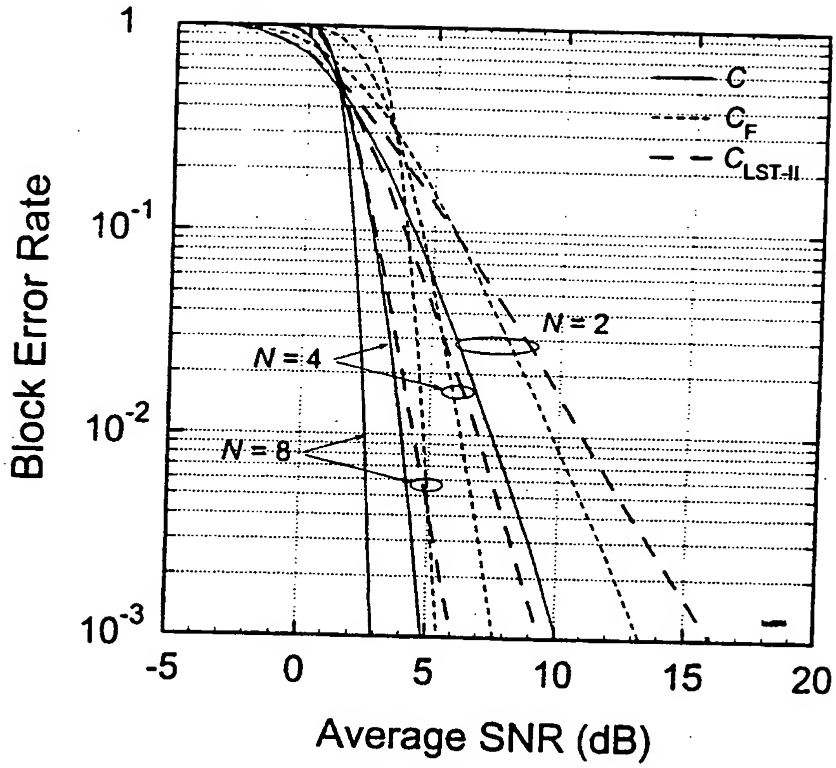


Fig. 4. Capacity bounds for quasi-static flat Rayleigh fading channels with N transmit and N receive antennas. C : Shannon capacity, C_F : Foschini (original) bound, and $C_{LST\text{-}II}$: capacity bound for LST-II.

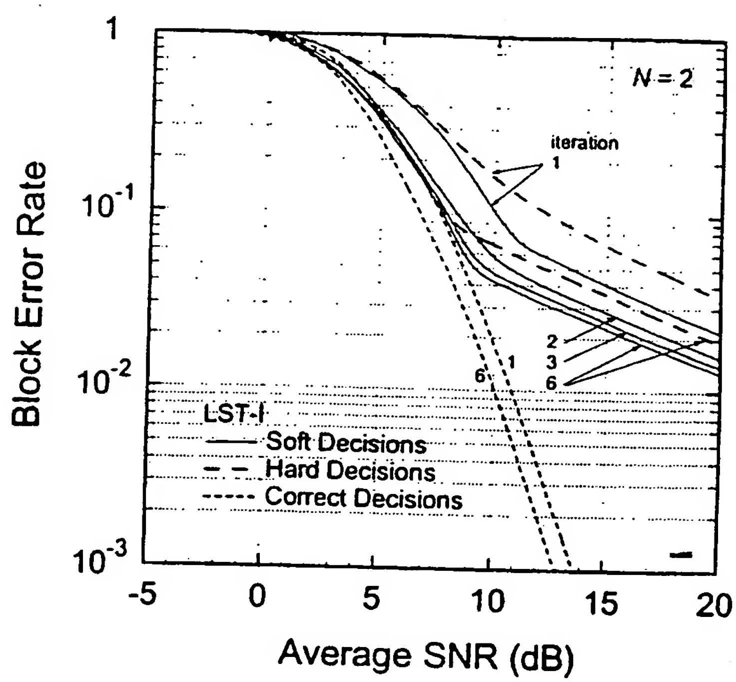


Fig. 5. Layered space-time performance of LST-I with 2 transmit and 2 receive antennas ($N = 2$). Quasi-static flat Rayleigh fading channel.

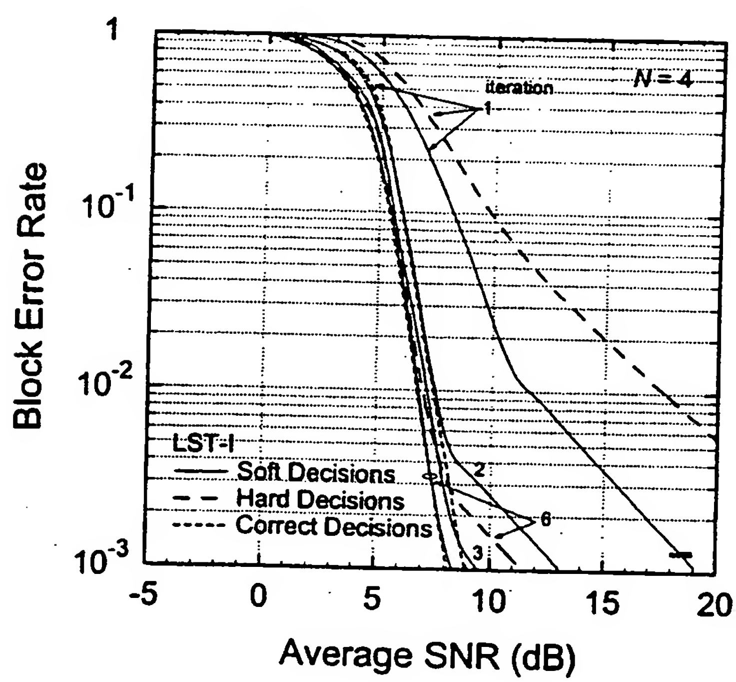


Fig. 6. Layered space-time performance of LST-I with 4 transmit and 4 receive antennas ($N = 4$). Quasi-static flat Rayleigh fading channel.

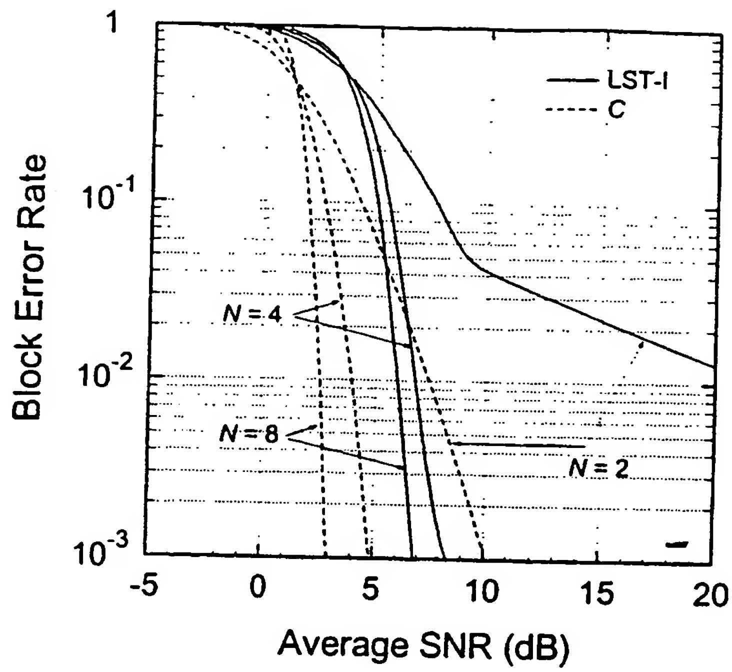


Fig. 7. Layered space-time performance of LST-I for $N = 2, 4$, and 8. Soft decisions. 6 iterations. Quasi-static flat Rayleigh fading channel.

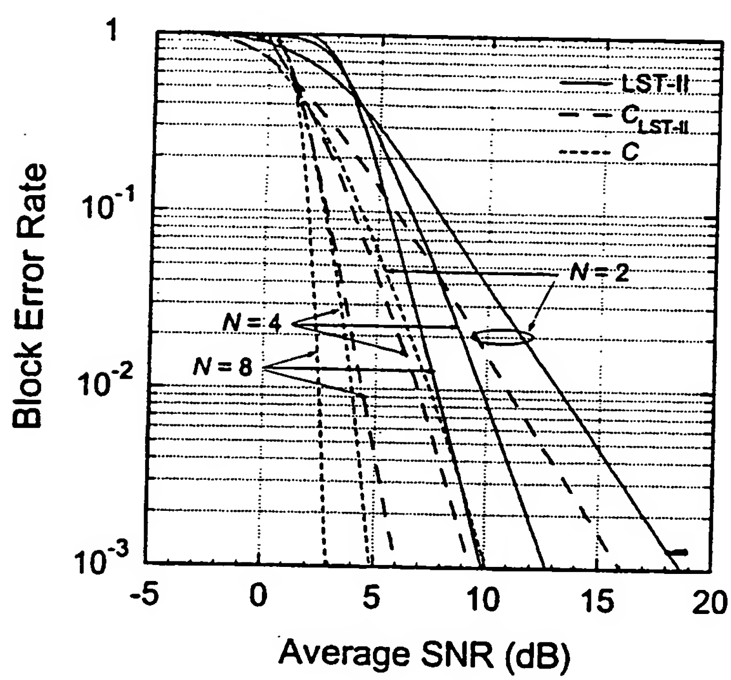


Fig. 8. Layered space-time performance of LST-II for $N = 2, 4$, and 8. Soft decisions. 2 iterations. Quasi-static flat Rayleigh fading channel.

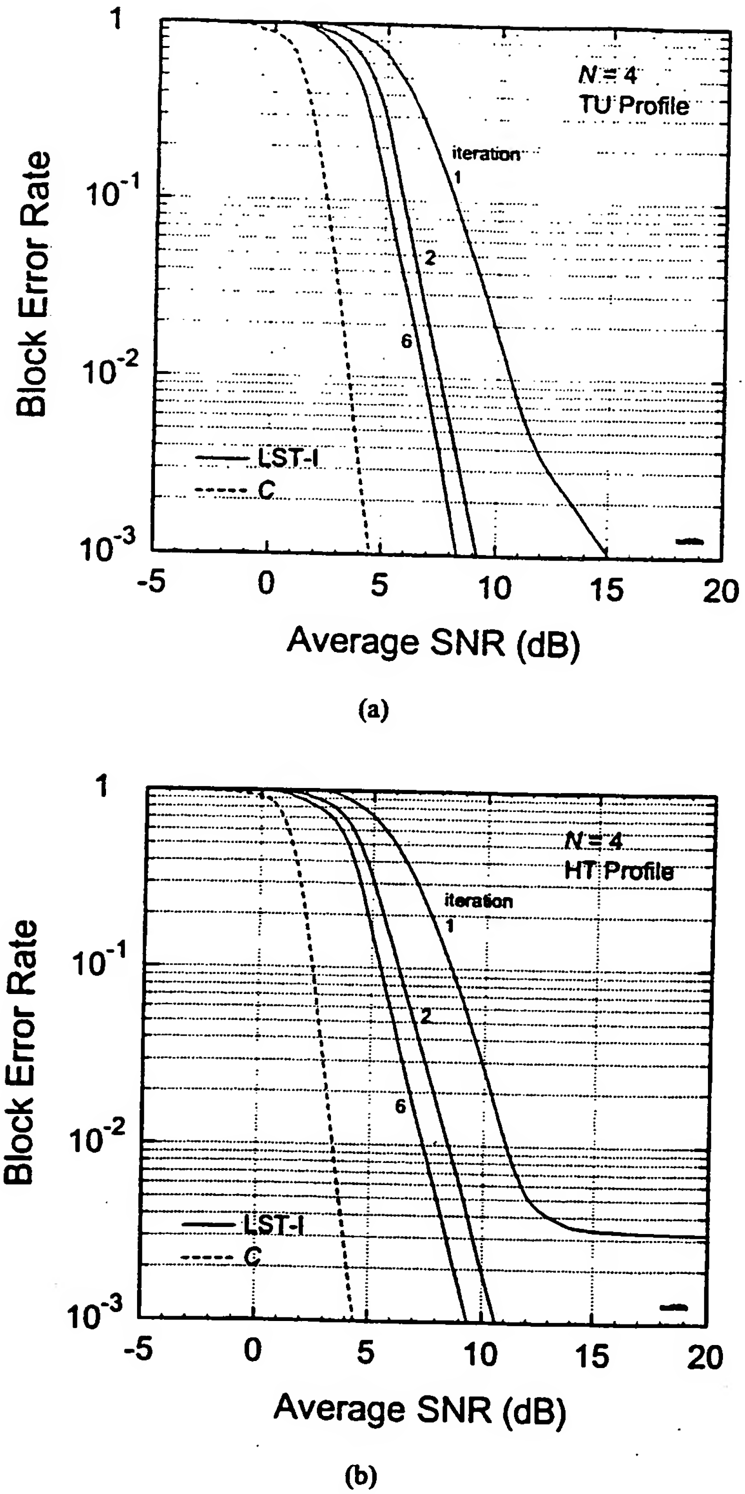


Fig. 9. Layered space-time performance of LST-I over frequency-selective channels: (a) TU profile. (b) HT profile. $N = 4$. Soft decisions.

INTERNATIONAL SEARCH REPORT

A. CLASSIFICATION OF SUBJECT MATTER
IPC 7 H04L1/06

According to International Patent Classification (IPC) or to both national classification and IPC

B. FIELDS SEARCHED

Minimum documentation searched (classification system followed by classification symbols)
IPC 7 H04L

Documentation searched other than minimum documentation to the extent that such documents are included in the fields searched

Electronic data base consulted during the international search (name of data base and, where practical, search terms used)

PAJ, WPI Data, EPO-Internal, INSPEC, COMPENDEX

C. DOCUMENTS CONSIDERED TO BE RELEVANT

Category °	Citation of document, with indication, where appropriate, of the relevant passages	Relevant to claim No.
X	DA-SHAN SHIU, KAHN: "Layered space-time codes for wireless communications using multiple transmit antennas" IEEE INTERNATIONAL CONFERENCE ON COMMUNICATIONS, 6 - 10 June 1999, pages 436-440, XP002152497 Piscataway, US page 437, left-hand column, paragraph 5 -right-hand column, paragraph 1 page 438, left-hand column, paragraph 2 -right-hand column, paragraph 1 ---- -/--	1-37

[X] Further documents are listed in the continuation of box C.

[ ] Patent family members are listed in annex.

° Special categories of cited documents:

"A" document defining the general state of the art which is not considered to be of particular relevance
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"T" later document published after the international filing date or priority date and not in conflict with the application but cited to understand the principle or theory underlying the invention
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"&" document member of the same patent family

Date of the actual completion of the international search	Date of mailing of the international search report
9 November 2000	30/11/2000

Name and mailing address of the ISA	Authorized officer
European Patent Office, P.B. 5818 Patentlaan 2 NL - 2280 HV Rijswijk Tel. (+31-70) 340-2040, Tx. 31 651 epo nl, Fax: (+31-70) 340-3016	Scriven, P

International Application No
PCT/US 00/24641

C.(Continuation) DOCUMENTS CONSIDERED TO BE RELEVANT

Category °	Citation of document, with indication, where appropriate, of the relevant passages	Relevant to claim No.
X	TAROKH ET AL.: "Combined array processing and space-time coding" IEEE TRANSACTIONS ON INFORMATION THEORY, vol. 45, no. 4, May 1999 (1999-05), pages 1121-1128, XP002152498 New York, US ISSN: 0018-9448 page 1125, left-hand column, paragraph 2 -column 5 ---	1-37
X,P	DA-SHAN SHIU, KAHN: "Scalable layered space-time codes for wireless communications: performance analysis and design criteria" IEEE WIRELESS COMMUNICATIONS AND NETWORKING CONFERENCE , 21 - 24 September 1999, pages 159-163, XP002152499 1999, Piscataway, NJ, USA, IEEE, USA ISBN: 0-7803-5668-3 the whole document ---	1-37
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